IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

APPLICANTS:	Muck et al.	SERIAL NO.:	10/531785
PATENT NO.:	7450490	FILED:	April 18, 2005
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ENTITLED: CHANNEL ESTIMATION USING THE GUARD INTERVAL OF A MULTICARRIER SIGNAL

REQUEST FOR A CERTIFICATE OF CORRECTION UNDER 37 CFR § 1.322

Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

Sir:

In accordance with the provisions of 37 CFR § 1.322 of the Rules of Practice, which implement 35 USC § 254, the Patent and Trademark Office is respectfully requested to issue a Certificate of Correction in the above-identified patent. It is certified that errors appear in the above-identified patent as shown in the attached Certificate of Correction. Applicant certifies that the errors are of a minor character and were not the fault of Applicant. Since the changes necessary to correct these errors in the patent would not constitute new matter, and would not require re-examination, Applicant prays a Certificate of Correction will issue. Since errors were not the fault of Applicant, it is believed that there will not be a fee for this Certificate of Correction. The Commissioner is further authorized to charge any additional fees that may be due, or credit any overpayment to Deposit Account **502117**.

Respectfully submitted,

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UNITED STATES PATENT AND TRADEMARK OFFICE
CERTIFICATE OF CORRECTION

PATENT NO.: 7450490

DATE: November 11, 2008

INVENTOR(S): Muck et al.

It is certified that error appears in the above-identified patent and that said Letters Patent are hereby corrected as shown below:

In Column 4, Line 58, delete "ak" insert - - αk - -, therefor.

In Column 7, Line 21, delete " $V_{HP}=[H_{(N+D)\times(N+D)}]\cdot(PD^{T}0_{N}^{T})^{T}$ " and insert - - $V_{HP}=\left[H_{(N+D)\times(N+D)}\right]\cdot\left(P_{D}^{T}0_{N}^{T}\right)^{T}$ - -, therefor.

In Column 11, Line 58, in Claim 1, delete "J" and insert - - j - -, therefor.

In Column 12, Lines 4 to 8, in Claim 1, delete "

$$[\hat{V}]=\left(\sum_{n=0}^{N+D-1}|\beta_k|-\frac{2n}{N+D}\right)^{-\frac{1}{2}}$$

$$diag\left\{1,\ \beta_k\frac{1}{N+D},\ \dots\ ,\ \beta_k\frac{N\_D-1}{N+D}\right\},$$

" and insert - -

$$[\hat{V}]=\left(\sum_{n=0}^{N+D-1}|\beta_k|^{\frac{2n}{N+D}}\right)^{-\frac{1}{2}}\bullet diag\left\{1,\beta_k^{\frac{1}{N+D}},\dots,\beta_k^{\frac{N+D-1}{N+D}}\right\}$$

- -, therefor.

In Column 12, Line 40, in Claim 1, delete " $S_{(F)}^{\hat{E}Q}(k)=[F_{NXN}]\cdot S^{EQ}$ " and insert - - $S_{F}^{EQ}(k)=\left[F_{NxN}\right]\cdot S^{EQ}$ - -, therefor.

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PATENT NO. 7450490

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FIG. **12** is a graph representing preferred values of prefixes used in the system of FIG. **1**.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

FIG. **1** shows an OFDM communication system in accordance with one embodiment of the invention comprising a transmitter comprising an OFDM modulator **1** and a receiver comprising an OFDM demodulator **2**, the transmitter and the receiver communicating over a communication channel **3**.

An input bit-stream $b_n \epsilon (0,1), n=0,1, \ldots, K-1$ is modulated onto a set of N carriers whose carrier amplitudes are given by the vector $X(k)=(X_0(k), X_1(k), \ldots, X_{N-1}(k))^T$, corresponding to OFDM symbol number k. Afterwards, the time domain OFDM signal is generated by means **4** which performs an Inverse Fourier Transform operation, or preferably an Inverse Fast Fourier Transform ('IFFT') operation $[F_N]^{-1}=[F_N]^H$ with $[F_N]^H=([F_N]^T)^*$ where $(\cdot)^T$ is the transposition operator and $(\cdot)^*$ is the complex conjugate operator:

$$x(k) = [F_N]^{-1} X(k) = (x_0(k), x_1(k), \ldots, x_{N-1}(k))^T., \qquad \text{Equation 1}$$

where

$$[F_N] = \frac{1}{\sqrt{N}} \cdot (W_N^{lk})_{0 \le l \le N-1, 0 \le k \le N-1},$$

and $W_N = e^{-j\frac{2\pi}{N}}$.

The resulting parallel signal x (k) vector is converted to a series signal by a parallel-to-series converter **5**, a prefix, represented by the D×1 vector $P_D=(c_0, \ldots, c_{D-1})^T$, being inserted into the signal as guard interval between each OFDM symbol to produce a series digital signal x_n. The series digital signal x_n is then converted to an analogue signal x(t) by a digital-to-analogue converter **6** and transmitted over the channel **3**.

The channel **3** has a Channel Impulse Response H(k)=C(k) and also introduces noise v.

At the receiver **2**, an analogue signal r(t) is received and converted to a digital signal r_n by an analogue-to-digital converter **7**. The digital signal r_n is then converted to a parallel signal by a series-to-parallel converter r(k) and equalised and demodulated by equalisation and demodulation means **9** to produce demodulated signals $s^{est}(k)$. In the following analysis, consideration of noise is omitted for the sake for simplicity. However, including the consideration of noise does not significantly modify the results.

In some known OFDM communication systems, the guard interval is used to add some redundancy (D samples of redundancy are added) by introducing a cyclic prefix, for example in the following manner:

$$x^{(CP)}(k)=(x_{N-D}(k), \ldots x_{N-1}(k), x_0(k), x_1(k), \ldots x_{N-1}(k))^T.$$

In other words, data from the end of the frame is repeated by the transmitter in the guard interval to produce a prefix.

In accordance with this embodiment of the present invention, however, the prefix samples inserted as guard interval of OFDM symbol number k, $\alpha_k.c_0$ to $\alpha_k.c_{D-1}$, are deterministic and are known to said receiver as well as to said transmitter. The prefixes comprise a vector $P_D=(c_0, \ldots, c_{D-1})^T$ of size D×1 that is common to the symbols multiplied by at least one weighting factor α_k, so that the prefixes have the overall form $\alpha_k.c_0$ to $\alpha_k.c_{D-1}$. The weighting factor α_k may be constant from one symbol to another. However, in a preferred embodiment of the invention, the weighting factor a_k differs from one symbol to another, the elements of a given vector P_D being multiplied by the same weighting factor. With an OFDM modulator in the transmitter functioning in this way, blind channel estimation in the receiver can be done simply and at low arithmetical complexity. In particular, the receiver can constantly estimate and track the channel impulse response without any loss of data bandwidth. Moreover, the demodulator at the receiver can have advantageous characteristics, ranging from very low arithmetical cost (at medium performance) to high arithmetical cost (very good system performance).

More particularly, in the preferred embodiment of the invention, the prefix of D samples that is added in the guard interval comprises a pre-calculated suitable vector $P_D=(c_0, \ldots, c_{D-1})^T$ of D samples that is independent of the data and that is weighted by a pseudo-random factor α_k that only depends on the number k of the latest OFDM symbol:

$$x^{(const)}(k)=(\alpha_k c_0, \ldots, \alpha_k c_{D-1}, x_0(k), x_1(k), \ldots x_{N-1}(k))^T. \qquad \text{Equation 2}$$

For the purposes of the analysis below, a second prefix/OFDM symbol vector is defined as follows:

$$x^{(const,post)}(k)=(x_0(k), x_1(k), \ldots x_{N-1}(k), \alpha_{k+1}c_0, \ldots, \alpha_{k+1}c_{D-1})^T. \qquad \text{Equation 3}$$

Several choices for α_k are possible. It is possible to choose $\alpha_k \epsilon C$, that is to say that α_k can be of any complex value. However, any α_k with $|\alpha_k| \neq 1$ leads to performance degradation compared to preferred embodiments of the invention.

It is possible to limit the choice of α_k, somewhat less generally to $\alpha_k \epsilon C$ with $|\alpha_k|=1$. This choice usually leads to good system performance, but the decoding process risks to be unnecessarily complex.

Accordingly, in the preferred embodiment of the present invention, the phase of α_k is chosen so that

$$\alpha_k = e^{j\frac{2\pi}{N+D} \cdot m},$$

where m is an integer, N is the useful OFDM symbol size and D is the size of the pseudo-random prefix. This choice is particularly advantageous when using the specific decoding methods described below.

For the sake of simplicity, the following analysis assumes that the weighting factor has been chosen as

$$\alpha_k = e^{j\frac{2\pi}{N+D} \cdot m},$$

m integer. However, it will be appreciated that the mathematical adaptation to any of the cases presented above is straightforward.

It proves to be very useful to choose $\boxed{a_k}$ such that its phase changes from OFDM symbol to OFDM symbol. The constant prefix P_D is preferably chosen with respect to-certain criteria, for example the following:

- In the frequency domain, P_D is as flat as possible over the frequency band used for data carriers.
- In the frequency domain, P_D is as near to zero as possible for all unused parts of the band.
- In the time domain, P_D has a low peak-to-average-power-ratio (PAPR).

Thus, this method works very well if the prefix-spectrum is non-zero everywhere in the $FFT_{D \times D}$ domain (and, of course, everywhere well above channel noise). This can be a troublesome limitation in other circumstances.

A second embodiment of a method of channel impulse response estimation on D carriers in accordance with the present invention avoids this limitation, at the expense of increased arithmetic cost. This second method does not estimate $\hat{h}_D$ based on a de-convolution in the $FFT_{D \times D}$ domain as presented above, but estimates $FFT_{(N+D)\times(N+D)}((\hat{h}_D{}^T 0_N{}^T)^T)$ directly based on the received vector $([H_0]+[H_1]) \cdot P_D$. This is possible by exploiting the observation:

$$[H_{(N+D)\times(N+D)}] \cdot (P_D{}^T 0_N{}^T)^T = (E_4{}^T E_0{}^T 0_{N-D}{}^T)^T \quad \text{Equation 13}$$

This equation is represented in more detail in FIG. **10**. In this second method, the channel impulse response is estimated using the following steps:

- Perform a $FFT_{(N+D)\times(N+D)}$ on $\boxed{V_{HP}=}$ $\boxed{[H_{(N+D)\times(N+D)}] \cdot (PD^T 0_N{}^T)^T}$
- Perform a $FFT_{(N+D)\times(N+D)}$ on $V_P = (P_D{}^T 0_N{}^T)^T$
- Perform a component-by-component division $\hat{H}_{N+D}{}^{(F)} = V_{HP} \oslash V_P$
- If desired, perform an IFFT on $\hat{H}_{(N+D)}{}^{(F)}$: $\hat{h}_{(N+D)} = IFFT_{(N+D)\times(N+D)}(\hat{H}_{N+D}{}^{(F)})$

The last step of the list presented above is not essential for the basic equalization algorithm but may be useful, for example in algorithms used to reduce noise levels.

The above methods have been described with reference to the specific case where α_k is constant and equal to 1. In preferred embodiments of the invention, however, the weight α_k of the prefix to each symbol k is a preferably complex pseudo-random factor that only depends on the number k of the latest OFDM symbol. The adaptations to this method of the basic equations (shown in FIG. **9**) are shown in FIG. **11**.

It is found that equations 4 and 8 are to be adapted as follows:

$$E_{\alpha,0} = \underbrace{E([H_0] \cdot r_0(k))}_{=0} + E\left([H_1] \cdot \frac{(\alpha_k \cdot P_D)}{\alpha_k}\right) = [H_1] \cdot P_D. \quad \text{Equation 14}$$

$$E_{\alpha,4} = \underbrace{E([H_1] \cdot r_3(k))}_{=0} + E\left([H_0] \cdot \frac{(\alpha_{k+1} \cdot P_D)}{\alpha_{k+1}}\right) = [H_0] \cdot P_D. \quad \text{Equation 15}$$

The procedures for blind channel estimation described above remain applicable by setting $E_0 = E_{\alpha,0}$ and $E_4 = E_{\alpha,4}$. This amounts to weighting the preceding and following D prefix-samples of each received symbol by the corresponding $\alpha_k{}^{-1}$ or $\alpha_{k+1}{}^{-1}$ respectively.

The values of the prefixes $\alpha_k \cdot P_D$ are chosen as a function of selected criteria, as mentioned above. Values that have been found to give good results with the criteria:

- Low Peak-to-Average-Power-Ratio of the time domain signal
- Low Out-of-Band Radiation, that is to say maximise the energy of the prefix over the useful band and not waste prefix energy over null carriers
- Spectral Flatness, e.g. SNR of each channel estimates shall be approx. constant
- Low-Complexity Channel Estimation, i.e. by prefix spectrum whose spectral contributions are mainly just phases (i.e. of constant modulus),

are shown in FIG. **12** by way of example, for the following OFDM parameters:

- Size of the Prefix in Time Domain: D=16 Samples
- Size of the OFDM symbols in the frame: N=64 Samples
- Carriers where channel coefficients are to be estimated (over N+D=80 carriers): Carriers 1 to 52
- Out-of-Band region: Carriers 76 to 80
- Maximum PAPR has not been limited
- Out-of-Band Radiation as low as possible
- Spectral Flatness as good as possible.

The channel estimation is done by calculating the expectation value over a number of samples of the received vector as explained above. If the tracking of the channel is done based on a first estimation $\hat{h}(k-1) = [F_D]^H \cdot \hat{H}(k-1)$ of the channel impulse response and a number R of OFDM symbols, the first estimate is then updated as follows:

$$\hat{H}(k) = s_0 \cdot \hat{H}(k-1) + \left([F_D] \cdot \sum_{n=0}^{R-1} s_{n+1} \cdot (E_0(k-n) + E_4(k-n))\right) \oslash ([F_D] \cdot P_D)$$

$$\hat{h}(k) = [F_D]^H \hat{H}(k).$$

based on the ideas of the first method for channel estimation that has been presented above. Alternatively, the second method can be applied by

$$\hat{H}(k) = s_0 \cdot \hat{H}(k-1) + \left([F_{N+D}] \cdot \sum_{n=0}^{R-1} s_{n+1} \cdot (E_4^T(k-n), E_0^T(k-n), 0_{N-D}^T)^T\right) \oslash \left([F_{N+D}] \cdot (P_D^T, 0_N^T)^T\right)$$

$$\hat{h}(k) = [F_{N+D}]^H \hat{H}(k).$$

where the factors s_n, n=0,1, . . . ,R−1 are positive real numbers that are used for normalization and weighting of the different contributions. Thus, for example it is possible to take older OFDM symbols less into account for the channel estimation than later ones. The Fourier matrix [F] can be chosen in the N+D carriers or D carriers domain.

Several equalization methods are advantageous using the pseudo-random prefix OFDM. In general, the different methods offer different performance-complexity trade-offs.

A first embodiment of a method of equalization uses zero forcing in the N+D Domain and offers low complexity equalization.

With

$$\beta_k = \frac{\alpha_k}{\alpha_{k+1}},$$

the Channel Impulse Response matrix can be represented as follows:

$$[H] = [H_{ISI}] + \frac{\alpha_k}{\alpha_{k+1}} \cdot [H_{IBI}] = [H_{ISI}] + \beta_k \cdot [H_{IBI}] \quad \text{Equation 16}$$

$$[H]\cdot\begin{bmatrix} I_N \\ 0_{D\times N}\end{bmatrix}:[G]=\left[[H]\cdot\begin{bmatrix} I_N \\ 0_{D\times N}\end{bmatrix}\right]^{+}.$$

Thus, the equalized resulting vector is

$$\begin{aligned} R^{(eq,ZF)}(k) &= [G]\cdot R^{(1)}(k) \\ &= [G]\cdot\left[[H]\cdot\begin{bmatrix} I_N \\ 0_{D\times N}\end{bmatrix}\cdot x(k)+v\right]. \end{aligned} \quad \text{Equation 20}$$

The definition of the Moore-Penrose pseudo-inverse is, among others, discussed by Haykin in the book: *"Adaptive Filter Theory"* by Simon Haykin, 3^{rd} edition, Prentice Hall Information and System Science Series, 1996. Haykin uses the common definition

$$[A]^{+}=(A^H A)^{-1}A^H. \quad \text{Equation 21}$$

where [A] is a rectangular matrix.

The invention claimed is:

1. A method of communication using Orthogonal Frequency Division Multiplexing ('OFDM"), the method comprising the steps of:

generating bit streams $b_n \in (0,1)$, n=0,1, . . . ,K−1 and the corresponding sets of frequency domain carrier amplitudes ($X_0(k)$ to $X_N(k)$), where k is the OFDM symbol number, modulated as OFDM symbols to be transmitted from a transmitter,

inserting prefixes as guard intervals in said sample streams,

transmitting said OFDM symbols from said transmitter to a receiver,

using information from said prefixes to estimate the Channel Impulse Response ($H_D^{(F)}$) of the transmission channels at the receiver, where ($H_D^{(F)}$) is the length D vector defined as the channel impulse response vector in the frequency domain, denoted by superscripted, F, and

using the estimated Channel Impulse Response ($\hat{H}_D^{(F)}$) to demodulate said bit streams in the signals received at said receiver, wherein said prefixes ($\alpha_{k.}\,c_0$ to $\alpha_{k.}\,c_{D-1}$) are deterministic and are known to said receiver as well as to said transmitter, where c is the set of vectors containing constant postfix samples, (α_k) is a weighting factor proportional to

$$e^{j\frac{2\pi}{N+D}m},$$

where $\boxed{J}$ is the square root of −1, N is the useful OFDM symbol size, D is the size of the prefix vector, and m is an integer, and further performing the multiplication by a matrix proportional to

$$R^{(1)}(k)=\sqrt{N+D}\cdot[\hat{V}]\cdot r(k),$$

where

-continued

$$\boxed{[\hat{V}]=\left(\sum_{n=0}^{N+D-1}|\beta_k|-\frac{2n}{N+D}\right)^{-\frac{1}{2}}\cdot diag\left\{1,\ \beta_k\frac{1}{N+D},\ \dots\ ,\ \beta_k\frac{\text{N\_D}-1}{N+D}\right\},}$$

where β_k is the ratio of consecutive α_k,

$$\beta_k=\frac{\alpha_k}{\alpha_{k+1}},$$

calculating the frequency shifted CIR coefficients

$$\hat{H}_{N+D}^{Shifted,F}=\begin{pmatrix} \hat{H}\left(\beta_k-\frac{1}{N+D}\right)\dots\ , \\ \hat{H}\left(\beta_k-\frac{1}{N+D}\cdot e^{j2\pi}\frac{N+D-1}{N+D}\right)\end{pmatrix},$$

performing a component-by-component division

$$R^{(2)}(k)=R^{(1)}(k)=\bigotimes\hat{H}_{N+D}^{Shifted,F},$$

performing a multiplication by a matrix proportional to

$$R^{(3)}(k)=[\hat{V}]^{-1}\cdot\frac{1}{\sqrt{N+D}}\cdot R^{(2)}(k),$$

extracting the N equalized samples corresponding to the k^{th} data symbol to the vector $S^{EQ}(k)$, and

transforming the symbol $\hat{s}(k)$ into frequency domain by performing a Fourier Transform: $\boxed{S_{(F)}{}^{EQ}(k)=[F_{\boxed{NXN}}]\cdot S^{EQ}}$ (k).

2. A method of communication as claimed in claim **1**, wherein said prefixes ($\alpha_{k.}c_0$ to $\alpha_{k.}c_{D-l}$) comprise a vector (P_D) that is common to said symbols multiplied by at least one weighting factor (α_k).

3. A method of communication as claimed in claim **2**, wherein said weighting factor (α_k) differs from one symbol to another but the elements of a given vector (P_D) are multiplied by the same weighting factor.

4. A method of communication as claimed in claim **3**, wherein said weighting factor (α_k) has a pseudo-random value.

5. A method of communication as claimed in claim **1**, wherein said weighting factor (α_k) is a complex value.

6. A method of communication as claimed in claim **5**, wherein the modulus of said weighting factor (α_k) is constant from one symbol to another.

7. A method of communication as claimed in claim **1**, wherein estimating said Channel Impulse Response

$$\left(H_D^{(F)}\right)$$

comprises performing a Fourier Transform on a first vector (V_{HP}) that comprises the received signal compo-